VOL. 23 No. 274

WIRELESS ENGINEER

JULY 1946

Editor
W. T. COCKING, M.I.E.E.

Managing Editor
HUGH S. POCOCK, M.I.E.E.

Technical Editor
Prof. G. W. O. HOWE, D.Sc., M.I.E.E.

Editorial Advisory Board.—F. M. COLEBROOK, B.Sc., A.C.G.I. (National Physical Laboratory), L. W. HAYES, O.B.E., M.I.E.E. (British Broadcasting Corporation), Professor E. B. MOULLIN, Sc.D., M.I.E.E., A. H. MUMFORD, O.B.E., B.Sc. (Eng.), M.I.E.E. (G.P.O. Engineering Department), R. L. SMITH-ROSE, D.Sc., Ph.D., M.I.E.E. (National Physical Laboratory).

EDITORIAL

Two Electromagnetic Problems

A LONG solenoid of small cross-section is wound on a non-magnetic core and supplied with a steady current I from a constant-current source. It is free to move in a magnetic field—say that of the earth—and takes up its normal position with its axis along the lines of force. If now we forcibly turn it round so that its north pole points to the south, we do work. To calculate the work done, consider one turn and assume that it is a square of side s; the force on each side is $sH_xI/10$ dynes where H_x is the strength of the field in which the solenoid is placed. The work done is turning the one turn through 180° will be $2s^2H_xI/10 = 2AH_xI/10$ ergs where A is the cross-sectional area of the coil. For T turns the work will be $2AH_xIT/10$ which, since $H = 4\pi IT/10l$ is equal to $HH_xAl/2\pi$ ergs, where H is the magnetic field produced within the solenoid by the current I. On turning the solenoid round the flux through it will be decreased by an amount $2AH_x$ and, since the current is maintained constant during the change, the energy supplied to the circuit will be $I\int edt = IT2AH_x/10 = HH_xAl/2\pi$ ergs.

Hence the mechanical work done in turning the coil is exactly equal to the electrical energy supplied by the coil to the circuit, but the resultant value of H within the solenoid has been reduced from $H + H_x$ to $H - H_x$ and the energy of the magnetic field within the solenoid has therefore been reduced and this appears at first sight to be unaccounted for. The amount of this reduction of energy is $[(H + H_x)^2 - (H - H_x)^2]Al/8\pi = 4HH_xAl/8\pi = HH_xAl/2\pi$ ergs, neglecting the small end-effects. Thus all the three amounts of energy concerned are equal. There is, however, a fourth amount of energy involved, since, when the solenoid is reversed, there is a strengthening of the external field, and the reduction of the magnetic energy within the coil is exactly balanced by the increase of magnetic energy outside the coil. At a great distance from the solenoid the field H_x is quite unaffected by the reversal and the total magnetic flux over a large cross-section is therefore unchanged; it is merely redistributed on reversing the coil.

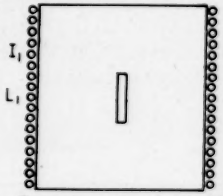

Fig. 1. *A small solenoid is located at the centre of a large cylindrical coil, L_1 carrying a current I_1.*

Imagine the field H_x to be produced by a large cylindrical coil, Fig. 1, with the small solenoid at its centre as shown and let the self-inductance of the large coil be L_1 and that of the solenoid L and let M be the mutual inductance. Let I_1 be the current

in the large coil producing the field H_x. With the solenoid in its normal position the magnetic energy will be $0.5\ L_1I_1^2 + 0.5\ LI^2 + MI_1I$. On reversing the solenoid, any tendency of the total magnetic flux linking the large coil to decrease can be counteracted by increasing I_1 to I_2 so that $I_1L_1 + IM = I_2L_1 - IM$ and therefore $I_2 = I_1 + 2IM/L_1$. The total magnetic energy will now be $0.5\ L_1I_2^2 + 0.5\ LI^2 - MI_2I$ which on substituting the above value for I_2 gives $0.5\ L_1I_1^2 + 2I^2M^2/L_1 + 2MI_1I + 0.5\ LI^2 - MI_1I - 2I^2M^2/L_1 = 0.5\ L_1I_1^2 + 0.5\ LI^2 + MI_1I$ which is the same as before.

It will be seen that the necessary increase of current $I_2 - I_1$ in the outer coil is proportional to M, and hence, as the coil is made larger in diameter and M correspondingly decreased, the necessity of changing the current ceases. If then the magnetic field is produced by something so far removed from the solenoid that it is unaffected by its reversal, there is no change in the energy of the magnetic field on reversing the solenoid, the decreased energy inside being balanced by the increased energy outside it. Hence, although one appears to be doing work against the forces of the magnetic field, the magnetic energy is merely redistributed and all the work done appears as energy in the electric circuit. Since, at distances large compared with the length of the solenoid, H falls off as the cube of the distance, the modifications of the magnetic field due to the reversal are very local.

Torque on a Solenoid

When the solenoid is held at right-angles to the direction of the field it is interesting to consider the nature of the torque exerted upon it. One's permanent magnet upbringing makes it natural to picture poles at each end upon which the magnetic field exerts forces in the direction of the field, but closer examination suggests that this is very far removed from the facts of the case.

Fig. 2 shows a current-carrying solenoid in a magnetic field H_x. In the absence of H_x the magnetic field of the solenoid itself causes forces on the turns, but these merely tend to make the turns expand and come closer together, without any tendency to make the solenoid rotate. When H_x is applied we need, therefore, only consider the forces on the turns due to their current flowing in the presence of the magnetic field H_x. These forces will be the same on each turn whether at the ends or at the middle of the coil and they will be parallel to the axis of the coil as indicated by the arrows. Hence the torque is not due to forces parallel to H_x acting on the ends of the solenoid but to forces at right-angles to H_x acting equally all along the solenoid. Assuming the turns to be square the force on each side is $sH_xI/10$ dynes, giving a torque of $s^2H_xI/10$ per turn and a total torque of $s^2H_xIT/10$. If m is the pole strength $4\pi m = 4\pi s^2IT/10l$ = total magnetic flux through the solenoid; hence torque $= mH_xl$.

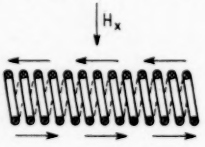

Fig. 2. *Current-carrying solenoid in magnetic field H_x. The arrow shows the direction of the forces on the turns.*

This shows that, although about half the flux never reaches the end of the solenoid but escapes sideways near the end, and although the poles are often said to be somewhere near the ends, the whole length must be taken in calculating the torque, that is, the poles must be assumed to be actually at the ends. The above discussion and Fig. 2 indicate, however, that the whole conception of the field acting on the poles, however convenient as a method of calculating the torque, bears little relation to the actual interaction between field and solenoid.

G. W. O. H.

PORTABLE PRECISION AMPLIFIER-DETECTOR*

By F. A. Peachey, A.M.I.E.E., S. D. Berry, and C. Gunn-Russell, B.A.

(Lines Department, British Broadcasting Corporation)

SUMMARY.—This paper is a description of a Portable Precision Amplifier-Detector, calibrated in decibels, for the measurement of the level of tone of 50 to 10,000 c/s from + 20 db to − 50 db to the nearest 0.1 db. Facilities for the measurement of programme volume and also for noise measurement down to − 110 db are incorporated.

THE amplifier-detector described in this article was designed primarily for precision testing of broadcast circuits and associated apparatus in the field. Routine measurements can be covered for most purposes by apparatus of less precision, but as this design has materialized it has become apparent that as so much extra may be accomplished for the addition of so little, precision gear may economically supplant the less precise type. This point is further emphasized by features in this detector which permit its separate use for normal or precision working. There are three uses, to any one of which it may be adapted by the operation of a single switch on the front panel. The same switch gives choice of high or low impedance input and its positions provide the following facilities :—

(1) Peak programme meter of 30,000 ohms input impedance.

(2) Peak programme meter of 600 ohms input impedance.

(3) Noise (peak value), 600 ohms input impedance.

(4) Level (calibrated R.M.S.), 600 ohms input impedance.

(5) Level (calibrated R.M.S.), 30,000 ohms input impedance.

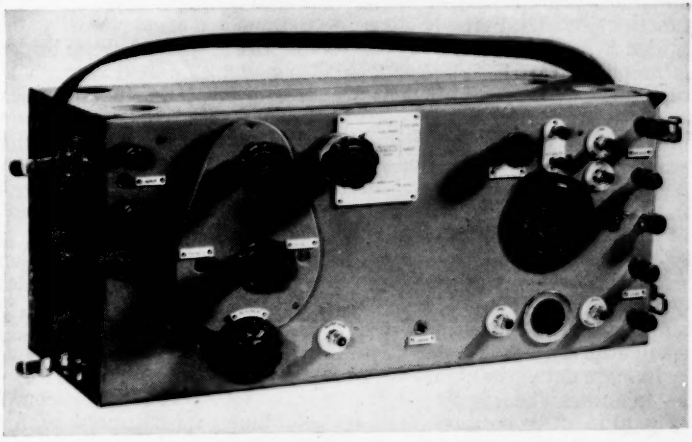

Fig. 1. *The complete amplifier-detector.*

The instrument (Figs. 1 and 7) is contained in a duralumin box 19in × 8¾in × 9in deep overall and weighs approximately 30 lb. It requires D.C. power supplies of 6 volts, 2 A and 200 volts, 12 mA. It has not been designed for supply direct from the mains, as it may frequently be used at places where such facilities are not immediately available. No doubt it would operate satisfactorily from a mains supply, although special steps would probably have to be taken to guard against hum induction from the valve filaments when it is used as a sensitive noise measuring device.

* MS. accepted by the Editor, January 1946.

Level Measurements

The circuit for level measurements will first be described as it covers the main part of the apparatus. Broadly, the circuit functions as follows (Fig. 2). A calibrated attenuator S_1R_2 followed by an amplifier A_1A_2 of known gain feeds a rectifier R, which supplies current to a D.C. meter M. The whole is calibrated so that a sinusoidal tone of given level across the input, deflects

the meter to a marked point when the attenuator switches are set to indicate that given level. This provides for measurements demanding an accuracy of ± 0.25 db. For greater accuracy an auxiliary gain control R_5 is provided whereby the output of the amplifier (simultaneously with the above function) is used to balance a thermal bridge B_1. The output of the thermal bridge is amplified by A_3, so that a small signal, representing lack of balance, is readily indicated, and the bridge may be brought into balance by adjustment of the auxiliary gain control R_5. Indication of the state of balance is given by a cathode-ray tuning indicator operating on the rectified output of the bridge amplifier A_3. The auxiliary gain control is such that a considerable angular movement produces relatively little change in the thermal-bridge currents and so may be calibrated quite widely to show small adjustments in level.

In more detail the circuit may be described as follows. The circuit selector switch S_2 (Fig. 3) is set to position (4) or (5) according to the input impedance required and connects the input terminals via the input transformer T_2 to the cascaded variable potential dividers S_1 and R_2. S_1 is calibrated in 10 db steps to cover a level range of + 20 to − 40 db, while R_2 is calibrated in 0.5 db steps with a range of 0 to − 9.5 db.* The output of these cascaded potential dividers supplies the input of a three-valve resistance-capacitance coupled amplifier. The overall negative feedback of the first two stages is controllable by R_5. This resistance operated over its full travel varies the gain of the amplifier by just over 1 db and is actually calibrated from + 0.5 to − 0.5 db. A further gain control, R_7, provides a gain adjustment for precision calibration purposes. The third and final stage to the amplifier supplies two output transformers, T_3 and T_6. The former provides negative feedback to V_3 and output for the thermal bridge. A 6-volt metal-filament lamp in one arm of the bridge reaches a resistance of 50 ohms at about 10 mA. At this value the bridge balances and no output is passed through V_4 and its output rectifier to the tuning indicator V_5. If the bridge is not balanced, the output of V_4 is rectified and operates the indicator. Critical balancing of the bridge is achieved by adjustment of, first, the coarse volume controls S_1, R_2 at the amplifier input and, secondly, the fine control, R_5. The gain of V_4 is such that a very clear indication is obtained if the bridge input current is changed 0.1 db from its balance value.

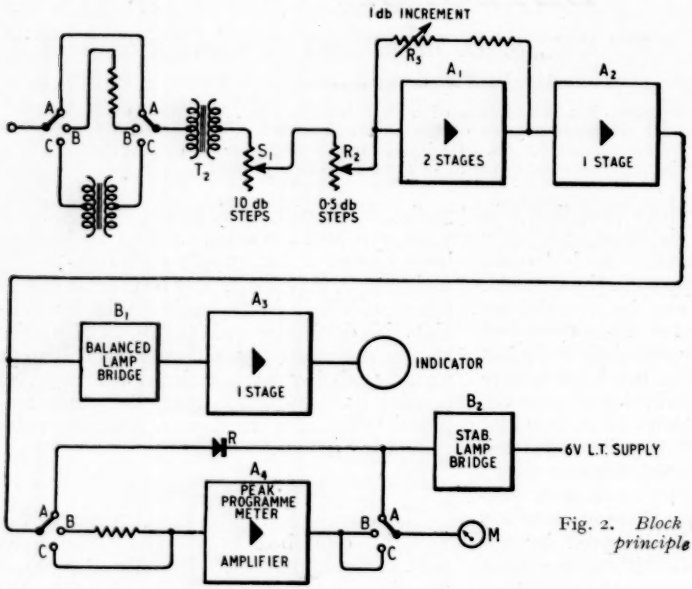

Fig. 2. *Block diagram illustrating the principle of the apparatus.*

Transformer T_6 provides, via the bridge metal rectifier, sufficient direct current to bring the meter to its calibrated point. The D.C. circuit is complicated by the fact that a special meter with a right-hand zero is used as this is necessary for the Peak Programme-measuring circuit described below. Most operators are used to a meter pointer moving in a clockwise direction, or left to right, as the

* Reference level 1 milliwatt in 600 ohms.

current is increased. To establish this feature a " backing-off " current is supplied to the meter in the reverse direction from that of the detector output and of a value such as to deflect the pointer to zero on the left-hand side of the scale with no signal from the rectifier. It is important that when this current has once been adjusted (by R_{13}) it should remain constant despite any normal

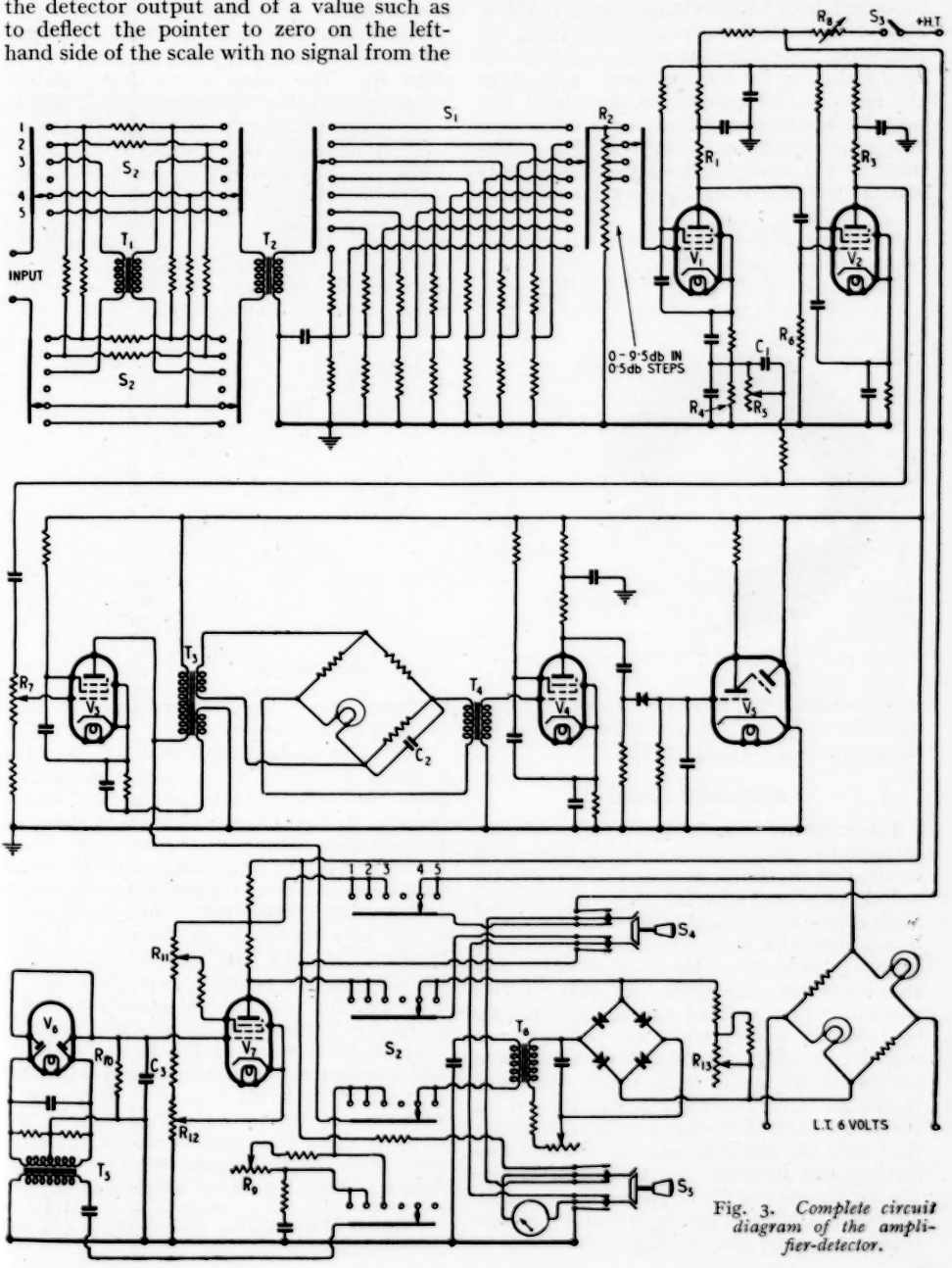

Fig. 3. *Complete circuit diagram of the amplifier-detector.*

variation in the L.T. supply. The thermal-stabilizing circuit shown ensures this. (See Appendix.) In using the gear, therefore, the procedure is to adjust the potential dividers S_1 and R_2 until the meter pointer is on (or nearly on) the calibrated mark, and read the level indicated by the switches. This gives the level to the nearest 0.5 db and it can actually be assessed rather more closely than this by interpolation. However, such assessment is not really necessary, as if greater accuracy is required, R_5 may be adjusted to show balance by means of the cathode-ray indicator. This gives the level reading to the nearest 0.1 db quite easily.

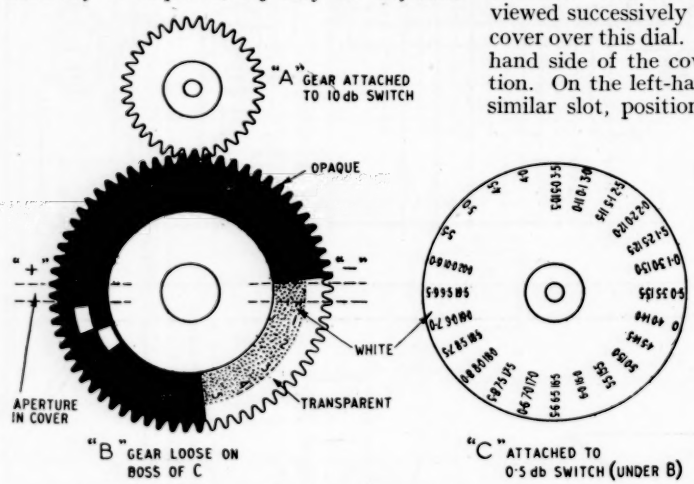

Fig. 4. *The main attenuator controls S_1 and R_2 have a combined indicator. "A" is attached to S_1 and "C" to R_2. "B" is the cover for "C" and has windows through which the "C" scale can be read. It is geared to "A" so that the operation of S_1 alters the position of the window and, therefore, the reading of the "C" dial.*

Attenuator Scales

Frequent use of such apparatus has drawn attention to an inaccuracy which may often occur, due to the incorrect addition of two switch readings, such as + 10 and − 5.5. Such errors are excusable when it is remembered that the user often has some more complex problem on his mind when he has reason to use the gear. As also in this particular case a third control (the 0.1 db control) is added with its positive and negative settings, it was felt that some assistance to avoid such errors was well justified. The two main controls, S_1 and R_2, therefore, have their dials so geared together that only the algebraic sum of the two dial settings can be seen. Furthermore, to emphasize the sign of the level, positive readings appear in a window on the left and negative readings appear on the right, the windows being clearly labelled " + " and " − ." The arrangement of this dial is shown in Fig. 4, whilst Fig. 5 shows the manner in which it is assembled.*

There are two main controls, the one providing 10 db steps S_1 and the other 0.5 db steps R_2. The range of the latter switch is 0 to 9.5 db. For clarity these will be called the tens and units switches respectively.

The unit switch drives dial " C " (Fig. 4). There are three concentric sets of figures on this dial. The inner figures 0 — 9.5 may be viewed successively through a slot in a fixed cover over this dial. This slot is on the right-hand side of the cover in a horizontal position. On the left-hand side of the cover is a similar slot, positioned so that both ranges of figures 0.5 — 10.0, and 10.5 — 20.0 may be seen (successively).

Fitted freely between dial " C " and the cover is a mask " B." This is driven by gear " A " which is attached to the " tens " switch. If the tens switch is set in the +20 position then the range of figures 10.5 — 20.0 is exposed in the left-hand slot ; the actual figure exposed being determined by the units switch setting (dial " C "). In a like manner if the tens switch is set to the " + 10 " position the mask " B " moves round covering all groups of figures except those 0.5 — 10.0 and the reading (determined by " C " setting) is displayed in the left-hand window. This left-hand window is labelled " Total + " and has figures exposed in it only when the tens switch is in its " + 20 " or " + 10 " position. With the " tens " switch in any other position the mask is moved round so that only the right-hand window displays figures. The right-hand window is labelled

* Details of the dial are the subject of a British patent application.

"Total —ve" and displays the outer set of units on dial "C" for all settings of the tens switch from 0 to —50. This is arranged by leaving the outer part of mask "B" transparent over an arc corresponding to this range (see Fig. 5). The appropriate

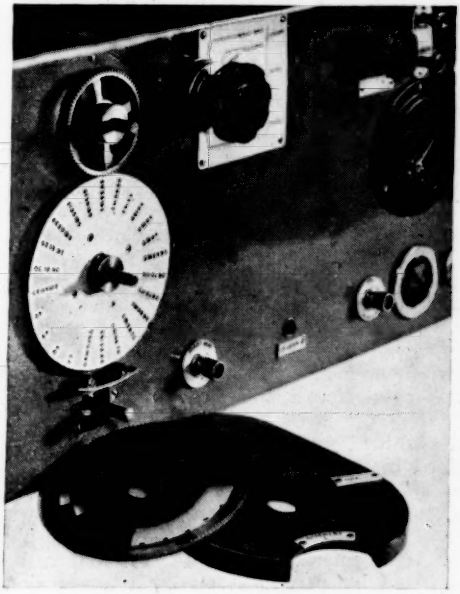

Fig. 5. *Details of the attenuator dial.*

"tens" digit is also carried by the mask and placed on the left-hand side of the units figure on "C" (Fig. 4). Actually, the whole arrangement is very simple and can be seen clearly from Fig. 5 which shows the "units" dial on the "units" switch shaft, the gear attached to the "tens" shaft and, lying in the foreground, the mask (normally driven by the tens switch) and the cover to expose one reading only on the left- or right-hand side. The units dial is constructed of white ivorine and marked with black figures. The mask is of perspex painted black for masking purposes, left clear where exposure of the units dial is required, and painted white to provide masking behind the tens digits and to match the ivorine dial underneath.

It should be observed that sundry small capacitances such as C_1 in the feedback path, C_2 in the bridge circuit, etc., have been employed in order to compensate for stray reactance effects and so provide for a flat response, or good degree of balance, at all frequencies. Table 1 shows that in spite of this it was impossible to arrange for a flat response at all levels. It was considered desirable to arrange for the best performance from + 5 to — 15 db in level, although it will be seen that even with an input level of — 55 db, the error is less than 0.1 db at 10,000 c/s.

The amount of negative feedback it was possible to provide in the amplifier is not as great as would be necessary for the degree of stabilization considered desirable. This may be seen from the following. It has been shown[1], employing the notation of the article referred to, where

A = highest gain of an amplifier without negative feedback.

TABLE 1

FREQUENCY RESPONSE

Voltage input to bridge balance.
Relative gain referred to zero at 1,000 c/s at various input levels.
Source approximately 600 ohms into "600-ohm level" input.
Probable error of method 0.02 db.

The first three columns show the effect of varying the 0.5 db input potential divider—the other columns the effect of the 10 db potential dividers. A significant error appears in the last few of the latter only.

Freq. c/s	Input level db.								
	0	— 5	— 9.5	+ 5	— 15	— 25	— 35	— 45	— 55
50	+ 0.02	+ 0.04	+ 0.02	0	0	+ 0.06	+ 0.04	+ 0.03	+ 0.04
100	+ 0.02								
250	+ 0.02								
500	0								
3,000	0								
5,000	+ 0.02	0	0	0	0	— 0.02	0	— 0.07	— 0.04
8,000	0								
10,000	0	0	— 0.02	0	0	— 0.06	— 0.04	— 0.15	— 0.07

A^1 = highest gain of the amplifier with negative feedback.

β = feedback ratio.

X = the fraction of its maximum value which the gain of the amplifier assumes when circuit constants are varied.

SF = stability factor.

$$= \frac{\text{percentage change of } A}{\text{percentage change of } A^1}$$

that $SF = 1 + A\beta X$.

In this case, the first stage A.C. load is R_1 in parallel with R_6 and equals 32,000 ohms, and that of the 2nd stage is R_3, the following grid resistance, and the resistance of the negative feedback path, all in parallel and equals 67,000 ohms. So that, assuming a g_m of 3 mA/V for each valve under working conditions, and with R_7 at maximum, for the first two stages.

$$A = 3 \times 0.032 \times 3 \times 0.067 \times 10^6$$
$$= 19{,}300 \text{ times}$$
$$= 85.7 \text{ db}$$

and, with the fine control, R_5 in its mid-position, since R_4 is 230 Ω and the remainder of the feed-back path 552,000 Ω.

$$\beta = \frac{230}{0.552 \times 10^6}$$
$$= 0.416 \times 10^{-3}$$

if the small effect of current feedback in the first stage from R_4 is ignored.

$$\text{Then } A^1 = \frac{19{,}300}{1 + (0.416 \times 10^{-3} \times 19{,}300)}$$
$$= 2{,}140 \text{ times}$$
$$= 66.6 \text{ db.}$$

giving a gain reduction due to negative feedback of 19.1 db. If X be taken as 0.8, a value below which it is not likely to fall in the circumstances of the use of the apparatus, then

$$SF = 1 + (19{,}300 \times 0.416 \times 10^{-3} \times 0.8)$$
$$= 7.4$$

So that the resulting change of gain under feedback conditions will be

$$\frac{20}{7.4} = 2.7 \text{ per cent} = 0.23 \text{ db} \quad .. \quad (1)$$

In the output stage the negative feedback is taken from a separate winding on the output transformer and is of such magnitude as to reduce the gain of the stage by 13 db.

For this stage, therefore

$$1 + A.\beta = 4.47$$
$$A.\beta = 3.47$$

and $$SF = 1 + (3.47 \times 0.8)$$
$$= 3.78$$

so that, with $X = 0.8$ as before, the resulting change of gain with negative feedback will be

$$\frac{20}{3.78} = 5.3 \text{ per cent} = 0.45 \text{ db} \quad .. \quad (2)$$

It was found that, under the conditions shown, the valves used changed their g_m by about 7 per cent for an anode voltage change of 20 per cent, the maximum that might occur in use. Calculated as above it will be seen that this would cause a gain alteration of nearly 0.3 db. To assist in guarding against changes of this order in the absolute calibration, a variable resistance R_8, suitably decoupled, is connected in the common anode lead. A push-button switch connects the indicating meter as a voltmeter across the H.T. supply and R_8 enables the latter to be adjusted to a predetermined value. With the H.T. voltage fixed in this manner and the valve heaters supplied from an accumulator the only cause of gain change of the amount shown in (1) and (2) above would be valve ageing or replacement. This can, of course, be avoided by periodical test and a selection of the spare valves to be carried with the instrument.

For precision measurements the input impedance of the instrument is of course an important factor. The variation shown in Table 2 is largely due to the input transformer which in order to save space and weight is smaller than it might be if better results in this direction were required. The instrument is, of course, calibrated to give a reading proportional to the voltage across the input terminals and no correction has been made for variations of input impedance.

Initially, it was assumed that the instrument would be used to read R.M.S. on a sinusoidal wave. It was realized, of course, that the deflection of the meter, although calibrated on a pure sine wave to read R.M.S., was actually proportional to the rectified mean value. For this reason, it was thought better to use a thermal device for the fine control so that the error in reading the meter would not be misleading if the form factor of the wave were other

than 1.11. Thus, although with an input of two tones of different frequency, but equal level, the coarse reading meter may read as much as 1 db low, balance on the fine control can only be achieved when all switches are set to read the level corresponding to the proper R.M.S. value of the resultant input signal.

TABLE 2
INPUT IMPEDANCE

Frequency c/s	Z ohms
High Z—"Level"	
50	18,500 /40°
100	22,000 /27°
250	25,500 /14°
1,000	27,200 / 5°
3,000	28,500 /2.5°
5,000	28,500 / 0°
10,000	29,000 / 0°
Low Z—"Level"	
50	590
250	597
1,000	599
10,000	599
High Z—"Volume"	
50	29,700
1,000	29,900
10,000	29,900
Low Z—"Volume"	
1,000	595
"Noise"	
1,000	530

Peak Programme Meter

For some years now, the British Broadcasting Corporation has standardized the method of measuring programme volume on a meter designed to register the full peak of the signal, if that peak be of 5 (or more) milliseconds in duration. In this instrument the deflection of the pointer is proportional to the logarithm of the input-signal voltage. In other words, the scale is divided into equal db increments and so, over reasonable limits, provides an indication of the loudness of the programme (Fig. 6). All stations are fitted with these instruments so that at any point the "volume" line up may be checked and watched during programme transmission. The term "volume" is used to distinguish measurements made on a fluctuating programme signal, from "level" measurements made on steady tone.

A comprehensive investigation some years ago indicated that in average programmes the peak value (ignoring peaks of less than 5 milliseconds duration) was about 8 db higher than the equivalent R.M.S. value of the programme. Thus, zero volume is defined as that which will peak 8 db higher than will zero level tone on a meter of the appropriate characteristics fed from the standard rectifier circuit. The meter is specially designed so that it gives full deflection on a signal lasting 5 milliseconds and this explains why the particular type of instrument shown in Fig. 6 is used. A larger meter would have been rather better for level indications but would not have been suitable for peak readings. It would, of course, be difficult to read the meter deflection if it were allowed to operate with the same speed in both directions, so the return movement is arranged to be relatively slow. Referring to Fig. 3, it will be seen that for volume measurements the circuit-selector switch, S_2, is set to position (1) or (2). In this position, apart from an additional input-loss pad, the calibrated input potential dividers, S_1 and R_2, followed by the three-valve amplifier, operate as for level measurements. The additional loss pad guards the amplifier against overload, as in "volume" measurements the peak value of the signal is at least 8 db higher than the peak value of the corresponding "level". Also, as will be seen from Fig. 6, the scale is so marked that it is possible to allow the meter to deflect 12 db higher than the normal balance point for level measurements. With switch

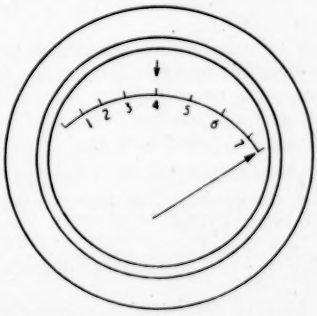

Fig. 6. *Scale of peak programme meter is 4 db per division between 1 and 7.*

S_2 in position (1) or (2) the output is switched from transformer T_6 to T_5. This latter

transformer provides the input to the double diode, the output of which rapidly charges C_3. If the input signal is removed (or diminished) C_3 discharges slowly through R_{10}. Thus the grid of the variable-mu pentode V_7 is energized suddenly by the rectified voltage of any peak signal and this voltage decays slowly. The time constant of charge is

$$T = CR$$

and since R is about 25,000 ohms

$$T = \frac{0.1 \times 25{,}000}{10^3} = 2.5 \text{ milliseconds.}$$

its characteristic such that the anode current is proportional to the logarithm of the input (grid voltage). The controls R_{12} (grid bias), R_{11} (screen-grid potential), and R_9 (sensitivity), are adjusted on test so that the correct law, sensitivity and zero reading with no input are arrived at. R_{12} and R_9 should not then need readjustment unless a valve is changed. R_{11} may be used as required to reset zero, if this is necessary due to normal variations in supply voltages, without disturbing the other requirements. All adjusters are arranged to operate by screw driver only

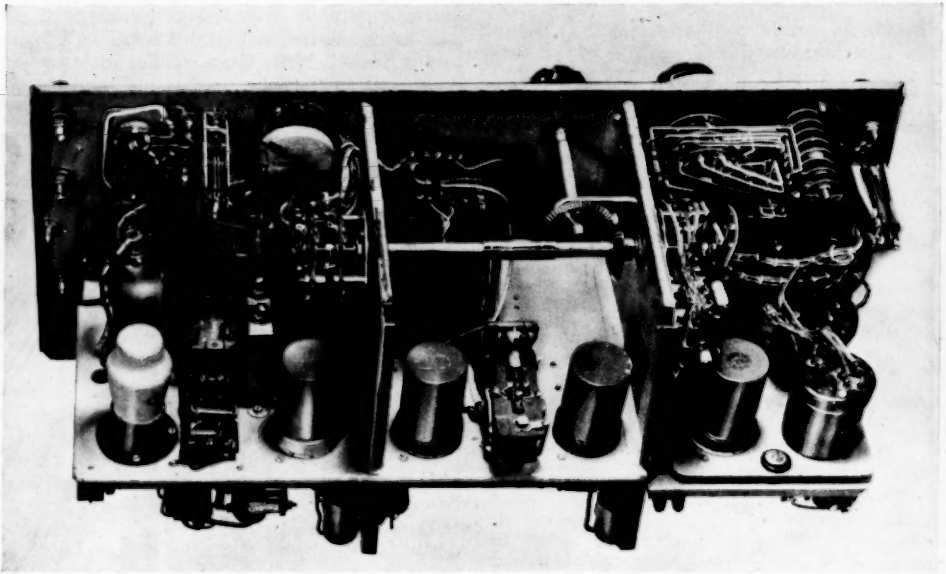

Fig. 7. *Interior of the amplifier-detector.*

With this value of time constant a pulse of E volts amplitude and 5 milliseconds duration will charge C_3 to a voltage E_1 such that

$$E_1/E = (1 - e^{-t/CR})$$
$$= 1 - e^{-5/2.5} = 0.865$$

that is C_3 will be charged to a voltage within about 1.2 db of E. The time constant of discharge is similarly

$$T = \frac{0.1 \times 10^7}{10^3} = 1{,}000 \text{ milliseconds.}$$

and the time of discharge is relatively slow, the capacitor voltage falling in 1 second to the fraction of its original value equal to e^{-1}, that is, nearly 9 db. The variable-mu pentode is arranged to operate on a part of

and are protected from the likelihood of accidental movement.

The amplitude—frequency response of the unit used as a peak programme meter is within about half a decibel from 50 to 10,000 c/s.

Noise (Peak Value)

It is logical to use the same type of instrument for the measurement of noise as has been used for the measurement of programme except that in the latter case a linear amplitude response is required, while in the former the response should be weighted to allow for the relative audibility of noise of different frequencies. Various curves have been

published from a vast accumulation of observations showing the variation in apparent loudness with pitch. From this information it would appear that the amount of weighting with respect to frequency depends on the general loudness of the noise from the absolute point of view. Rather than fit this piece of apparatus with weighting networks of this kind, it was decided to keep its response reasonably flat and connect any required networks externally. This arrangement left the instrument fairly simple for its size and afforded a facility for actual noise and crosstalk measurements at very low levels indeed.

One of its main uses is met at new broadcasting points where at an early stage no amplifiers or even power supplies are available. The land line is to be tested in advance to ensure that it is satisfactory for programme transmission and to determine values of equalization. Obviously, it would be erroneous to assess the noise on such a circuit before its equalizer is installed. In the course of these tests, therefore, a temporary (at least) equalizer is constructed and a noise measurement made. Now the equalized transmission equivalent of such a circuit may easily be 40 to 50 db. For cross talk and other reasons the sending level is not often above + 10 db so that a satisfactory noise level in some cases should have an absolute level not exceeding — 100 db. The maximum sensitivity of this device is therefore arranged to be 110 db.

This is accomplished by some sacrifice of frequency response. When switch S_2 is set to position (3) an extra input transformer replaces the fixed input loss pad described earlier, and the double diode receives a greater part of the output from V_3. This has the total effect of increasing the gain from input to meter by 40 db, so that a signal of — 90 db will provide a deflection to " 6 " and an input of — 110 db a deflection of " 1." Due largely to the big step-up in voltage provided by the input transformer (impedance ratio 1 : 50) the inherent noise level of the unit is such that no deflection can be observed on the meter when there is no input signal. The use of such a transformer prohibits a high-impedance input with maximum sensitivity, but so far no need has been felt for this. The response for peak noise measurements is shown in Table 3.

In conclusion, the authors wish to emphasize that in the individual circuits they can claim nothing new, but that the combination of these circuits has resulted in a compact and accurate amplifier-detector incorporating other useful facilities.

TABLE 3

FREQUENCY RESPONSE

Volume		Noise	
Frequency c/s	db	Frequency c/s	db
50	− 0.2	50	− 2.7
250	− 0.1	250	− 0.5
1,000	0	1,000	0
5,000	+ 0.1	5,000	+ 0.3
10,000	0	10,000	+ 0.1

The authors wish to express their gratitude to the British Broadcasting Corporation for their permission to publish this article, and to Messrs. Rantzen and Rendall for general help and advice during the development of the apparatus.

Reference

[1] Becker : " Negative Feedback in Amplifiers ". *Proc. Instn. Rad. Engn.*, June 1944.

APPENDIX

The resistance of a lamp alters considerably with temperature and the relation between the input current and output voltage of a bridge connected as in Fig. 8 is of the form shown in Fig. 9. At the

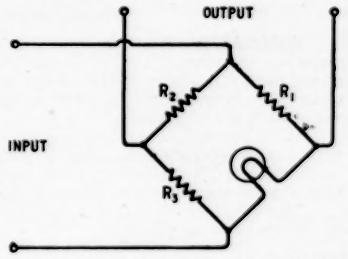

Fig. 8. *Basic circuit of a lamp bridge.*

point C the current through the lamp is of such magnitude as to cause the lamp to have a resistance equal to $\frac{R_1R_3}{R_2}$ thus balancing the bridge. By using a sensitive device to indicate this balance the bridge can be made to measure with accuracy a given input power level. This form of lamp bridge is used as a standard of level measurement in the Precision Amplifier-Detector. The lamp is an ordinary 6-V indicator lamp; $R_2 = R_3 = 2{,}000$ ohms and $R_1 = 50$ ohms.

With these values the bridge will balance at an input power of about + 10db, varying somewhat with the particular lamp in use. It has been found that lamps intended for use in a bridge of this kind must be aged by burning them at their rated voltage

for at least eight hours, after which they become stable. It has also been found that the balance point of the bridge is affected by the ambient temperature to an extent which falls as the lamp working temperature is increased, and that the power input to the bridge should not be, for this reason, much below + 10 db at which level the balance point alters by 0.25 db for an ambient temperature change of 45° F. The small capacitor

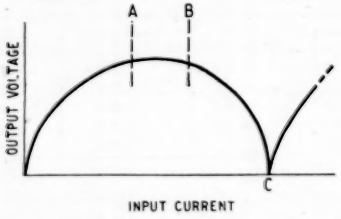

Fig. 9. *Relation between input current and output voltage of a lamp bridge.*

connected across one arm of the bridge in the Precision Amplifier Detector is to balance out the small reactance of the resistance cards which otherwise causes an ill-defined balance at high frequencies.

The region between the points A and B of Fig. 9 shows that over this range of input current the output voltage of the bridge remains practically constant due to the fact that the tendency of the output to rise, because of the rising input, is counteracted by the tendency to fall, due to the approach to balance. It can be shown that—with R_2 replaced by another lamp—when the current taken by the output circuit is small compared with that taken by the lamp and the resistance—current curve of the lamp is a straight line over the range required, generally approximately true, the percentage variation in output is $-\frac{D^2}{100}$, where D is the percentage variation in input.

A stabilizing bridge of this kind is used in the Precision Amplifier-Detector to stabilize the current required to bias-back the peak programme meter for use in the level-measuring circuit, so that variations in the L.T. supply, from which the current is derived, do not affect appreciably the zero setting of the meter. The lamps are the same type as before and R_1 and R_3 are each 150 ohms. A resistance in series with the input is adjusted so that the stabilization occurs over the required range (5.5 V to 6.3 V) of battery voltage.

POWER PULSE GENERATOR*

By M. Levy, A.M.I.E.E., M.Brit.I.R.E.

(Research Laboratories of the General Electric Company)

SUMMARY.—Description of a highly efficient pulse generator† in which voltage-pulse amplification is avoided by producing directly pulses of the required amplitude in a high impedance. The power of this initial generator is very small but it is amplified by a succession of cathode-followers which increase the peak power, lower the impedance and do not affect very much the amplitude.

With two S.T.C. valves type 4071A working as a pulse generator, and one S.T.C. valve type 4033A working as a cathode-follower, one can obtain pulses of about 1,600 volts in an impedance of 1,500 ohms and with two more 4033A valves in parallel, working as a second cathode-follower, one gets about 1,000 volts in 150 ohms.

The efficiency of such a circuit (output power divided by power supplied to the circuit) is of the order of 20 per cent.

THE use of pulse transmitters has been developed considerably since 1939. The present pulse generator was developed at that date as a modulator for a power-pulse transmitter. The requirements were to produce pulses of 1,000 volts in an impedance of about 150 ohms at a repetition frequency of 50 cycles per second. The peak power was about 6.5 kilowatts and the mean power about 50 watts.

To obtain this power, the usual method was, at that time, to use a low-power pulse generator followed by a pulse amplifier. The disadvantage of this solution is its inefficiency because in every other valve in the amplifier circuit current must flow all the time with the exception of the small intervals when the pulses are produced.

To eliminate this loss of power, the writer tried to avoid voltage amplification and to use valves in which the current flows only when the pulses are produced. To obtain this result, he used a low-power but high-voltage pulse generator, followed by peak-power amplifier valves, instead of voltage-amplifier valves, all working at nearly the same pulse amplitude, that is cathode-followers mounted in series. The pulse generator has a low

* MS. accepted by the Editor, February 1946.

† British Patent pending.

efficiency but, since it is only a lower-power generator, it does not greatly affect the overall efficiency. On the other hand, the power amplifier valves have a high efficiency and improve considerably the overall value.

In order to obtain pulses of large amplitude and of low mean power, low-power valves are used with voltages much greater than the

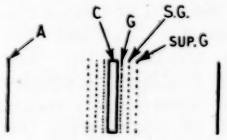

Fig. 1. *Diagrammatic section of a high-slope pentode. The grid, screen and suppressor are very close to the cathode, the anode is much further away.*

rated values applied to the anode, for instance 1,500 volts instead of 200 to 300 volts. This is, of course, a very unusual method, but with high-slope pentodes the perturbations produced are reduced to a minimum. In these valves, the grid, the screen and the suppressor, are very close to the cathode while the anode is at a distance. Fig. 1 shows diagrammatically a 5A-105A Standard valve.* The suppressor is about 1.5 millimetres from the cathode and the anode about 5 millimetres. If the grid, screen, and suppressor are working at the rated values, a high voltage on the anode will not affect the electrostatic field near the cathode and deterioration of the cathode by ion bombardment will not occur. The writer used this type of valve with an anode voltage of 1,500 volts without any apparent deterioration of the characteristics. However, the tests lasted only some days and due to the urgency of war requirements, no methodical study has been made of this question.

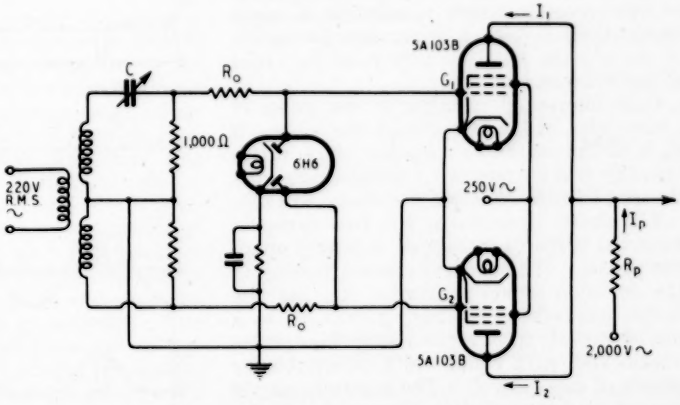

Fig. 2. *Circuit of a high-voltage low-power pulse generator. Large pulse amplitude is obtained by applying a high supply voltage to the anode circuit of the pentodes.*

* The 5A-105A Standard valve is nearly equivalent to the R.C.A. television high-slope pentode type 1852.

1. High-voltage Low-power Pulse Generator

For simplicity let us start from the simplified circuit of Fig. 2. The primary of a transformer is fed by a sinusoidal source having the same frequency as the required recurrent frequency. The secondary feeds the grids of two pentodes in nearly opposite phase, the approximation being adjustable by a small capacitor C. Each winding gives a high peak voltage of the order of 1,500 volts applied to the grid through a high dropping resistance R_0. The positive voltage on the grid is limited by the grid current. The negative voltage is limited in amplitude by a diode. The anodes of the two pentodes are connected together and fed by an H.T. supply voltage of the order of 1,000 volts or more.

Fig. 3 shows how the pulses are produced in this circuit. In (*a*) are represented the two sinusoidal voltages applied to the grids of the pentodes through high resistances. These waves are not exactly in opposition of phase, the shift being adjusted by the capacitor C. In (*b*) is reproduced in dotted lines one of these sinusoidal voltages. As its amplitude is very great, only a small fraction is transmitted to the grid, the difference being a drop of voltage across resistance R_0 produced by the grid and diode currents. The voltage really applied on the grid is represented by the heavy line. It has the shape of a succession of trapezoids placed alternately above and below the time axis. The slope of the edges is usually very great. With a peak amplitude of 1,500 volts for the sinusoidal wave, and high-slope pentodes, such as the

Standard valve type 5A.103B (cut-off at −9 volts), the time of build up is of the order of one thousandth of the duration of each trapezoid. With a recurrence frequency of

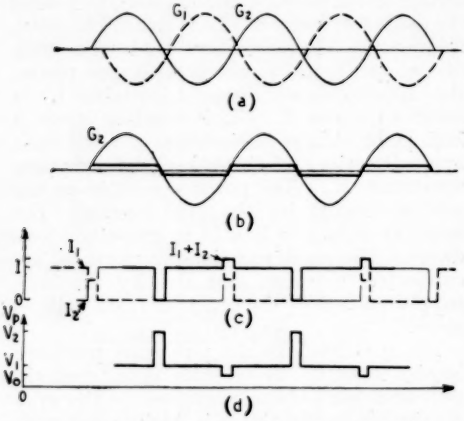

Fig. 3. *Diagrams explaining how the pulse generator of Fig. 2 works when a high-supply voltage is used*:— (*a*) *sinusoidal voltages applied on the grids* G_1 *and* G_2 *of the pentodes through resistances* R_0; (*b*) *sinusoidal voltage applied to grid* G_1 *through resistance* R_0 *and effective voltage appearing on this grid*; (*c*) *anode currents* I_1 *and* I_2 *and resultant current* $(I_1 + I_2)$; (*d*) *output pulses appearing across resistance* R_p.

50 c/s this gives a time of build-up of about 10 microseconds, which is sufficient in many applications. With a recurrence frequency of 500 c/s, the build-up time is of the order of one microsecond.

Peak sinusoidal voltages of the order of 1,500 volts will not overload the valves if R_0 is of the order of 1 megohm, since in this case the grid current is 1.5 milliamperes and the grid voltage does not rise above 10 volts.

The anode currents of the two pentodes have the shape of rectangles in nearly opposite phase. The resultant current flowing in the common anode resistance R_p, has the shape represented in Fig. 3 (*c*). It is a succession of positive and negative pulses interlocked, with equal width adjustable by means of capacitor C. The resultant anode voltage is represented in Fig. 3 (*d*). If the H.T. voltage applied to resistance R_p is 2,000 volts, the pulses have an amplitude not much smaller if R_p has a sufficiently high value.

When current is flowing in the valve, the anode voltage falls because of the voltage drop in resistance R_p. If this resistance has a high value, the drop of voltage can be sufficiently great to make the power dissipated at the anode very low.

This power can be still further reduced by feeding the anodes and screens with a sinusoidal H.T. voltage of frequency equal to the pulse recurrence frequency. The working diagram of the circuit is shown in Fig. 4 where (*a*) and (*b*) represent respectively the voltage applied to the grid of one of the pentodes through resistance R_0 and the current in the common anode circuit. The sinusoidal H.T. voltage feeding the anodes through resistance R_p is shown at (*c*) and is approximately 90° out of phase with the sinusoidal voltages applied on the grid circuits and is maximum when the anode current is zero. The anode voltage has the shape represented in (*d*). Where the anode voltage differs from the H.T. supply voltage, the latter is represented by dotted lines.

The anode voltage is maximum or negative when there is no current and is low when current flows. For instance, if the anode current is 200 milliamperes, with an H.T.

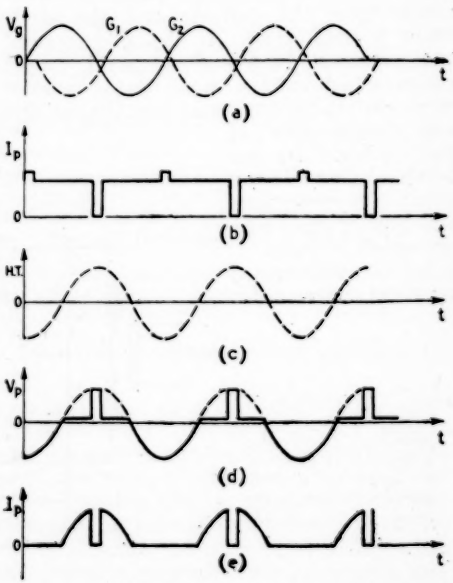

Fig. 4. *Diagrams explaining how the pulse generator of Fig. 2 works when an A.C. high-tension supply voltage is used*; (*a*) *same as Fig.* 3 (*a*); (*b*) *same as Fig.* 3 (*d*); (*c*) *sinusoidal H.T. voltage supply applied on the anodes of the pentodes through resistance* R_p; (*d*) *output anode voltage*; (*e*) *current flowing through resistance* R_p.

peak voltage of 2,000 volts and a resistance R_p of 10,000 ohms, the anode voltage when current flows is less than 100 volts, while the peak pulse voltage is nearly 2,000 volts. Each valve works during 1/4 of a period and the mean current is $\frac{I}{4\sqrt{2}}$ or about 1/6 of the peak current. The peak current can thus be 6 times greater than the average rated value. Also, since the anode voltage is one-half to one-third of the normal value, this peak current can be 2 or 3 times greater still, that is 12 to 18 times the normal rated current.

The screens are also fed by a sinusoidal source in phase with the source supplying the anodes, but the peak voltage must not be too much greater than the normal screen rated voltage. For instance, on the screen of valves type 5A.103B the sinusoidal peak voltage must be less than 300 volts (the rated value is 250 volts). In this case the cathode, the grid, the screen and the suppressor are working under normal conditions.

Fig. 5 shows a circuit with two pentodes type 5A.103B or 4071A. The normal anode current of these valves is 50 mA and the anode power dissipation is 12 watts. The valves are used with a peak current of 300 mA which leaves a good margin of safety. The anode is supplied with a sinusoidal wave of 2,000 volts peak value, and R_p is 7,000 ohms. The transformer gives two secondary voltages of 1,000 volts R.M.S. each, corresponding to a build-up time of 10 to 20 microseconds according to the type of valves used. The width of the pulses is adjustable between some tens to some thousands of microseconds.

2. Power-Pulse Amplifier

The pulse generator described above gives pulses of high voltage but low power, in a high resistance. We must now add to this generator a power-pulse amplifier which does not affect too much the pulse amplitude but increases the peak power. Such a power amplifier is really an impedance transformer and cathode-followers give exactly the results required.

A single-stage circuit is represented in Fig. 6. The pulses are applied to the grid and the output is taken from a resistance connecting the cathode to earth. The pulse amplitude is transferred from grid to cathode with a very small loss, while the cathode resistance can be much smaller than the input impedance. For example, with a 4033A (triode of 25 watts, slope 9 mA/V, filament consumption 8.5 watts), working with 2,000 volts applied to the anode, one can get 1,500 volts peak pulse amplitude on the cathode across a resistance of 1,500 ohms with a grid current of 50 mA, the grid being fed by the circuit of Fig. 5, that is by pulses of 2,000 volts peak amplitude appearing across a resistance of 7,000 ohms. Current in the valve flows only during the generation

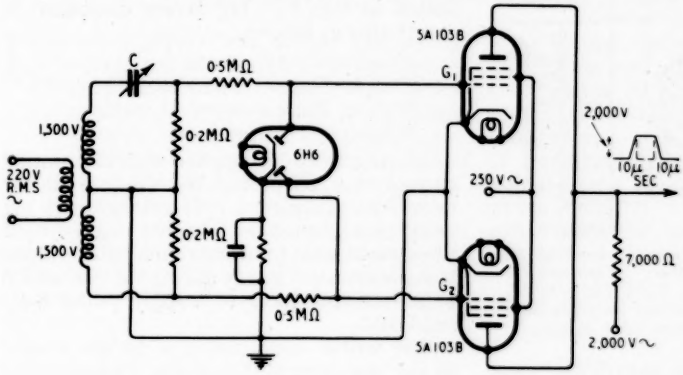

Fig. 5. *Practical circuit of the high-voltage low-power pulse generator producing pulses at a recurrence frequency of 50 c/s. The pentodes are fed directly by the mains through a convenient low-time-constant transformer.*

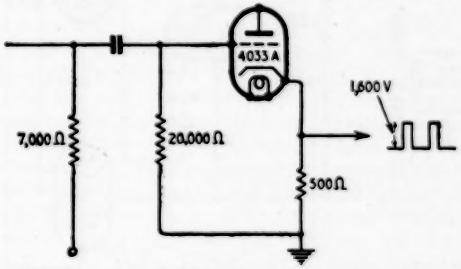

Fig. 6. *Cathode follower working as a peak-pulse power amplifier.*

of pulses. When a pulse appears, the voltage drop between anode and cathode is 500 volts for pulses of 1,500 volts. Hence if the valve

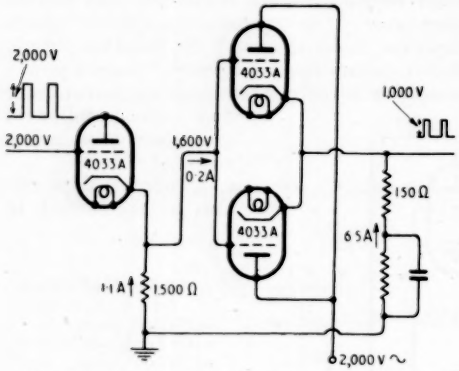

Fig. 7. *Two-stage cathode-follower working as a peak-pulse power amplifier. When coupled to the pulse generator of Fig. 5 it produces pulses of about 1,000 volts in an impedance of 150 ohms (6.5 kilowatts peak power).*

dissipates 25 watts, the maximum mean pulse power is 75 watts, which represents an efficiency of 75 per cent.

To get pulses of greater peak power, one can use many cathode-followers in series. Such a circuit is represented in Fig. 7. The first stage is a cathode-follower with a resistance between cathode and earth of 1,500 ohms or more. The second stage has two 4033A valves in parallel and a resistance of 150 ohms connecting the cathodes to earth. For a grid current of 350 milliamperes the cathode current is of the order of 6.5 amperes. This stage gives pulses of 1,000 volts and its efficiency is 50 per cent. This means that the output mean pulse power is 50 watts maximum, that is, smaller than for the preceding stage, *but the peak pulse power has increased from* 1.6 *to* 6.5 *kilowatts.*

The circuit of the pulse generator followed by the peak-power pulse amplifier is represented in Fig. 8. The power dissipated is distributed as follows :

Generator : 150 watts.
Output Stages : 250 watts.
Output Pulse Power : 50 watts.
Efficiency : 12 per cent.

All stages can be supplied with alternating current, the frequency being equal to the recurrence frequency. If transformers are used, care should be taken to reduce their time constants to a minimum since these transformers work only during the generation of pulses and have to supply power very suddenly.

The writer constructed a circuit similar to the one represented in Fig. 8 generating pulses at a recurrence frequency of 50 c/s. Three small 75-watt transformers were used to supply power to the generator and the peak-power amplifier. The whole circuit was of very small dimensions and was used to modulate a R.F. valve of 200 watts.

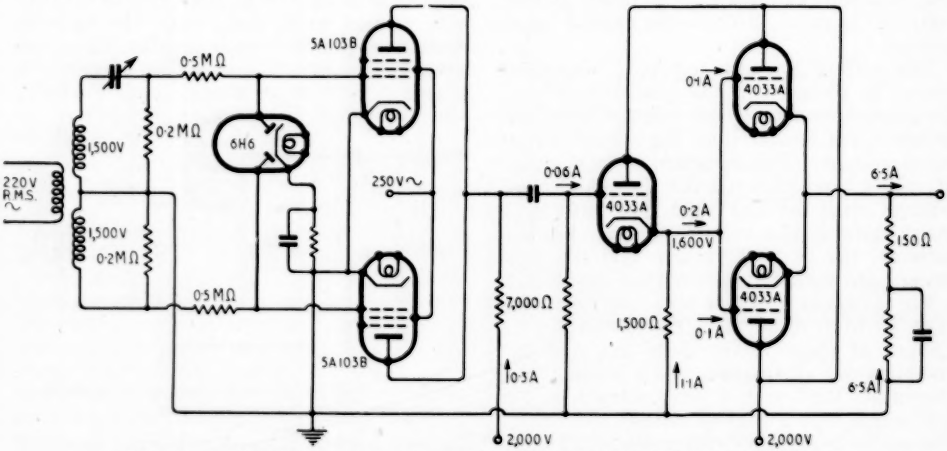

Fig. 8. *Peak-pulse power generator with an output of* 6.5 *kW peak-pulse power.*

This study was made in the "L.M.T. Laboratories" in Paris in 1939. Acknowledgements are due to Standard Telephones and Cables for permission to publish these results.

FILTER DESIGN TABLES BASED ON PREFERRED NUMBERS*

High-Pass Filters

By H. Jefferson, B.A., A.Inst.P., A.M.I.E.E.

IN previous papers† the author has shown how preferred numbers can be applied to filter design so that the capacitance values are necessarily preferred, and has given tables for the design of constant-k low-pass filters. In this paper tables are given for the design of constant-k high-pass filters. As before, the values in the tables are those of the ladder network elements. Where single sections only are used the actual principal element values are those shown in Fig. 1; L_k and C_k are the values obtained from the tables, and are approximations to

$$L_k = \frac{R}{4\pi f_c}$$

$$C_k = \frac{1}{4\pi f_c R}$$

where R is the design impedance and f_c is the cut-off frequency.

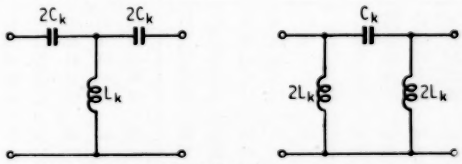

Fig. 1. *Basic T- and π-sections.*

For all preferred values of X there exist preferred values of $2X$, and these are listed below for reference.

As before, the tables are in pairs, each pair being used to determine the value of one element of the filter. The first table of each pair is of coarse mesh, in half-decade (logarithmic) steps. From this, the order of magnitude of the capacitance or inductance is found by examination of the values which border it. The second table covers only one decade and enables the significant figures to be found.

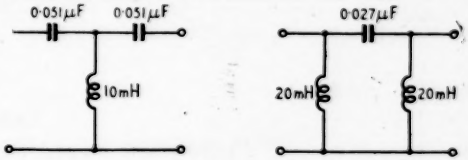

Fig. 2. *Single sections of a filter having the values derived in the text are shown here.*

As an example of the use of the tables, let us consider a high-pass filter of 620 ohms impedance and 5,100 c/s cut-off frequency. From Table I, the capacitance is in the square bounded by $8.2 \times 10^{-2}\,\mu$F, $2.7 \times 10^{-2}\,\mu$F, $8.2 \times 10^{-3}\,\mu$F and $2.7 \times 10^{-2}\,\mu$F. From Table III, the inductance is in the square bounded by 27 mH, 8.2 mH, 2.7 mH. and 8.2 mH. Table II gives the significant figures for the capacitance as 27, so that the actual value of capacitance is 0.027 μF. Table IV gives the significant figures for the inductance as 10, so that the actual value of the inductance is 10 mH. An exact computation gives $L_k = 9.6$ mH and $C_k = 0.0251$ μF. The physical form of a single section is shown in Fig. 2.

* MS accepted by the Editor, January 1946.

† "Preferred Numbers and Filter Design": *Wireless Engineer*, October 1945. "Filter Design Tables Based on Preferred Numbers — Low-pass Filters": *Wireless Engineer*, January 1946.

PREFERRED VALUES OF $2X$ FOR PREFERRED VALUES OF X.

X	10	11	12	13	15	16	18	20	22	24	27	30	33
$2X$	20	22	24	27	30	33	36	39	43	47	51	56	62

X	36	39	43	47	51	56	62	68	75	82	91	100
$2X$	68	75	82	91	100	110	120	130	150	160	180	200

TABLE I

CAPACITANCE C_k OF HIGH-PASS FILTER

DESIGN IMPEDANCE (OHMS)

CUT-OFF FREQUENCY	1	3.3	10	33	100	330	1 000	3 300	10 000	33 000	100 000	330 000
1	82×10^{3}	27×10^{3}	8 200	2 700	820	270	82	27	8.2	2.7	0.82	0.27
3.3	27×10^{3}	8 200	2 700	820	270	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}
10	8 200	2 700	820	270	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}
33	2 700	820	270	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}
100	820	270	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}
330	270	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820
1 000	82	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270
3 300	27	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270	82
10 000	8.2	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270	82	27
33 000	2.7	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270	82	27	8.2
100 000	0.82	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270	82	27	8.2	2.7
330 000	0.27	8.2×10^{-2}	2.7×10^{-2}	8.2×10^{-3}	2.7×10^{-3}	820	270	82	27	8.2	2.7	0.82
c/s	μF					$\mu\mu$F						
Mc/s	$\mu\mu$F					$\mu\mu\mu$F						

TABLE II

SIGNIFICANT FIGURES FOR C_k OF HIGH-PASS FILTER

DESIGN IMPEDANCE

CUT-OFF FREQUENCY	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	100
10	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82
11	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75
12	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68
13	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62
15	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56
16	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51
18	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47
20	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43
22	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39
24	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36
27	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33
30	30	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30
33	27	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27
36	24	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24
39	22	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22
43	20	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20
47	18	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18
51	16	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16
56	15	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15
62	13	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13
68	12	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12
75	11	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11
82	10	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10
91	91	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91
100	82	75	68	62	56	51	47	43	39	36	33	30	27	24	22	20	18	16	15	13	12	11	10	91	82

TABLE III

INDUCTANCE L_k OF HIGH-PASS FILTER

DESIGN IMPEDANCE—OHMS

CUT-OFF FREQUENCY	1	3.3	10	33	100	330	1 000	3 300	10 000	33 000	100 000	330 000
1	82	0.27	0.82	2.7	8.2	27	82	270	820	2 700	8 200	27 000
3.3	27	82	0.27	0.82	2.7	8.2	27	82	270	820	2 700	8 200
10	8.2	27	82	0.27	0.82	2.7	8.2	27	82	270	820	2 700
33	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82	270	820
100	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82	270
330	0.27	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82
1 000	82	0.27	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27
3 300	27	82	0.27	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2
10 000	8.2	27	82	0.27	0.82	2.7	8.2	27	82	0.27	0.82	2.7
33 000	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82	0.27	0.82
100 000	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82	0.27
330 000	0.27	0.82	2.7	8.2	27	82	0.27	0.82	2.7	8.2	27	82
c/s	μH						mH					
kc/s	μH x 10^{-3}						μH					
Mc/s	$\mu\mu$H						μH$\times 10^{-3}$					

TABLE IV

SIGNIFICANT FIGURES FOR INDUCTANCE L_k OF HIGH-PASS FILTER

DESIGN IMPEDANCE

CUT-OFF FREQUENCY	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	100
10	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82
11	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75
12	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68
13	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62
15	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56
16	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51
18	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47
20	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43
22	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39
24	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36
27	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33
30	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27	30
33	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24	27
36	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22	24
39	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20	22
43	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18	20
47	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16	18
51	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15	16
56	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13	15
62	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12	13
68	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11	12
75	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10	11
82	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91	10
91	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82	91
100	82	91	10	11	12	13	15	16	18	20	22	24	27	30	33	36	39	43	47	51	56	62	68	75	82

NEW BOOKS

Heaviside's Electric Circuit Theory

By H. J. Josephs, M.I.E.E. Pp. 115 + viii and 15 Figs. Methuen & Co., Ltd., London. Price 4s. 6d.

This is one of the series of monographs on physical subjects. The author is senior physicist at the Post Office Engineering Department and Dr. W. G. Radley, the Controller of Research, has contributed a foreword in which he says that the book was written at the suggestion of Professor Willis Jackson with the needs of electrical engineering students in mind. As Dr. Radley says, circuits are becoming more complicated and the growing importance of transient effects makes problems harder to solve. When he refers to the Heaviside Operational Calculus as being, in effect, a simple shorthand technique for evaluating the results of Fourier integral analysis, he is evidently using the word " simple " in a relative sense.

An outstanding feature of the monograph is the development of a " far-reaching theorem " which the author has reconstructed from the scattered papers of Heaviside ; as it was probably the last theorem that Heaviside deduced, the author has called it " Heaviside's Last Theorem." The first chapter is entitled " The Foundations of Electric Circuit Theory " ; in stating Kirchhoff's Laws, the author changes the order from that usually employed, and calls the circuit law the first and the currents meeting at a point the second law. Moreover, he takes liberties with the circuit law and states it as follows : " In any closed circuit in a network of conductors the algebraic sum of the potential differences is zero." In other words, if you return to the point from which you started the ups and the downs exactly cancel out, which is very true and obvious, but entirely misses the point of Kirchhoff's law.

Clerk Maxwell* says " Kirchhoff has stated the conditions of a linear system in the following manner, *in which the consideration of the potential is avoided.* (1) (Condition of " continuity "). At any point of the system the sum of all the currents which flow towards that point is zero.

(2) In a complete circuit formed by the conductors the sum of the electromotive forces taken round the circuit is equal to the sum of the products of the currents in each conductor multiplied by the resistance of that conductor."

Throughout this chapter the author has made a brave attempt to write " capacitance " and not " capacity " ; occasionally he succeeds, but very rarely. The second chapter deals with the expansion theorem. The author has obviously made a study not only of the Heaviside theorems, but also of the mental processes by which Heaviside arrived at many of his results and it is interesting to trace the paths which he followed. The expansion theorem is further developed in the third chapter which deals with power series solutions.

Chapter 4 deals with ladder networks, artificial lines and cables. Here again, when he comes down from mathematics to electrical matters, the author appears to be careless of details ; Fig. 3 which is stated to be composed of a series of T-sections, is incorrectly drawn at the receiving end, and the same applies to Fig. 4. The description is also confused, for he speaks of the middle of a section when the end of the T-section is meant ; the confusion is apparently between T- and Π-sections. Chapter V deals with impulse functions, transfer operators, and spotting functions and discusses the contributions of Carson and Bromwich to the subject. Chapter VI is entitled " The Establishment of Operational Formulae " and includes the " shifting " transformation and fractional order derivatives. Chapter VII is devoted to the very practical subject of transmission lines, one of the most important applications of Heaviside's methods. The final chapter is concerned with the application of modern theories of integration to the solution of circuit problems and deals with complex integration (something is missing in line 5 on p. 95), contour integrals, expansion theorems and Fourier integrals.

This monograph is a valuable addition to the Heaviside literature. It can be heartily recommended to the more mathematically minded students of electrical engineering. G. W. O. H.

* " Elementary Treatise on Electricity."

Plastics for Electrical and Radio Engineers.

By W. J. Tucker and R. S. Roberts. Pp. 148 + xi. Published by The Technical Press Ltd., " Piccancot," Gloucester Rd., Kingston Hill, Surrey, Price 12s.

The introductory chapter on " The Chemistry of Plastics and Insulating Materials " covers the subject matter briefly and forms a necessary introduction to a subject which has advanced in recent years to such an extent that it is an unknown quantity to many engineers. It is felt that this chapter might well have been expanded to link plastics with the materials with which they will be associated under service conditions, i.e. liquid dielectrics and metals. There is no mention of the catalytic action of metals and the electrolytic effects on the insulation value of plastics with catalysts under high electrical stress, where ionization, gas occlusion and ageing effects accumulate to deteriorate the efficiency of the material as a dielectric, often to ultimate breakdown.

It is interesting to note that some of the polymers of the liquid types are added to insulating oils as viscosity adjusters and to improve the electrical properties of the latter, particularly in respect of power factor, dielectric resistance and reduction in tendency to produce gas or polymerization products at high voltages.

This aspect of the problem will be of most interest to the electrical engineer, but it is of considerable importance to realize the breadth of the field covered by the substances generically termed " plastics." It would be of value if the authors could consider, for possible future editions, an expansion of the chemical section to include allied dielectric materials such as hydrocarbon oils and synthetic oils of the silicone type, particularly as many plastics are derived from petroleum products.

The second chapter forms an excellent survey

of the field of " Materials Available " and makes a valuable guide to the choice of materials in the design stage.

Eight tables are included which give all the essential information on the mechanical and electrical characteristics of phenol, urea-formaldehyde and aniline derived materials together with thermoplastics, synthetic rubbers, silicone and melamine plastics.

The treatment of silicones merits expansion in view of their increasing importance. Their products range from low-viscosity fluids for use in transformers, switchgear and similar applications to high-polymer resins for impregnation processes.

A chapter on terms and definitions follows and is well conceived ; it includes standard specifications. There is, perhaps, a bias in the direction of high-frequency applications, but with certain qualifications, which are referred to in the book, the fundamental principles apply equally at power-supply frequency.

The authors then turn to the important matter of the choice of material to suit the design requirements.

The allocation of articles into one or other of five classes according to functional requirements enables the designer to decide quickly the generic class of plastic for a specific purpose.

This is followed by a chapter on moulding and manufacturing processes, which contains a comprehensive resumé of the many possible methods of moulding plastics, and practical information on suitable cements. It is undoubtedly of assistance to all engineers, who are faced with repair problems involving plastic materials.

This chapter, together with the succeeding chapter VI, should assist designers to avoid some of the glaring examples of faulty mechanical or electrical design which have been perpetrated because of a lack of adequate appreciation of the medium.

The remainder of the book is devoted to specialized applications, the coating of wires to serve as an insulating covering, the production of insulated sleeving, insulation impregnants and cable coverings, with particular reference to the polyvinyl-chloride, polythene and polystyrene groups.

The book as a whole is an excellent example of a technical data book on a complex subject, for the presentation is acceptable and it is accurate in matters of fact. It can be recommended as a practical reference book to designers, engineers and students whose work necessitates essential data for applications in the electrical and radio fields.

G. W. I.

Mechanische Eigenschaften quasi-elastischer isotroper Körper

By FRIEDRICH POPERT. Pp. 105, and 30 Figs. Verlag A.G. Gebr. Leemann & Co., Zürich.

This paper-covered monograph, with the subtitle " Development of a new method of measurement for determining the shear modulus and internal friction in connection with a new process of television large scale projection " is a thesis presented at the Technische Hochschule at Zürich. The work was done under the direction of Professor Fischer and a number of references are given to papers by him on the subject ; these would have to be consulted before one could really get a clear idea of the object of the research. Apparently, the fluorescent screen of the cathode-ray tube is replaced by a thin layer of transparent semi-conducting material. This layer is known as the Eidophor. This material undergoes deformation which must disappear in the period between two successive pictures, *i.e.*, in 1/25 second. This led to the examination by experimental methods of the mechanical properties of suitable materials. The first half of the book is devoted to the theory of deformation and relaxation of elastic, viscous, plastic and quasi-elastic materials. In the second half the experiments are described for determining the modulus and the internal friction when subjected to small sinusoidal deformations at frequencies of from 10 to 55 cycles per second. These mechanical properties were determined from the electrical behaviour of the material such as variation of capacitance, and this part of the work involved elaborate circuit arrangements with modulators, demodulators, low-pass filters, valve voltmeters, etc. Diagrams and photographs of the apparatus are given and the results are analysed and plotted.

G. W. O. H.

Radar

By Major R. W. Hallows, T.D., M.A.Cantab., A.M.I.E.E. A simple non-technical explanation of the basic principles of radiolocation. Pp. 140. Chapman and Hall, Ltd., 37, Essex Street, London, W.C.2. Price 7s. 6d.

Basic Mathematics for Radio Students covers elementary algebra, geometry and trigonometry before proceeding to treat complex numbers, series, vectors, operators and the differential and integral calculus. It is intended primarily for radio engineers, but the main stress is on the mathematical ideas and it is thus equally suitable for students of other subjects. The application of mathematics to radio is chiefly dealt with in the concluding chapter.

Written by F. M. Colebrook, B.Sc., D.I.C., A.C.G.I., the book aims at helping students whose mathematical education has been interrupted by the war and throughout the method is to discuss basic ideas in some detail rather than to lay down rules of thumb and formulae. It is published by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 10s. 6d. There are 270 + x pages.

Radio-Communication Convention

THE Institution of Electrical Engineers is holding a Radio-Communication Convention on March 25th-29th, 1947. Like the recent radar convention, its aim is to present papers covering wartime wireless development, but this time in the field of communications.

The annual dinner of the Radio Section of the Institution will be held on the first day of the convention.

CORRESPONDENCE

Letters of technical interest are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Power Loss in Electromagnetic Screens

To the Editor, "Wireless Engineer."

SIR,—In the article by Davidson, Looser and Simmonds, appearing in your January issue, the eddy current density in the infinite shielding plane is derived theoretically for a single wire loop, while the measurements were made by using a coil, of length comparable to its distance from the screen, in order to energize this screen.

It is quite easy to extend the method of the authors of the above article in order to obtain an expression for the eddy current density when the energizing factor is a coil, instead of a single loop, and I should like to do so here:

Let the coil be situated so that its axis is perpendicular to the infinite shielding plane and the distance of its middle plane from the shield is a, while its length is $2c$. Using the same notation as in the above-mentioned article and the same units, one can just as easily show that the magnetic potential of the coil is given now by the expression

$$A_\phi' = (\mu/\pi)\, bnI \cos \omega t \int_{a-c}^{a+c} \int_0^{\pi/2} \frac{2 \sin^2 \theta - 1}{[(b+\rho)^2 + (z-Z)^2 - 4b\rho \sin^2 \theta]^{1/2}}\, d\theta dZ$$

where I is the current per turn, n the number of turns per unit length, and Z the longitudinal coordinate of a point situated on the coil.

The magnetic potential of the eddy currents is, then,

$$A_\phi = (\mu/\pi)\, bnI\omega \int_0^{\infty} \int_{a-c}^{a+c} \int_0^{\pi/2} \frac{(2 \sin^2\theta - 1)\sin \omega(t-\tau)}{[(b+\rho)^2 + (Z - z - \frac{2s}{\mu}\tau)^2 - 4b\rho \sin^2 \theta]^{1/2}}\, d\theta dZ d\tau$$

and, through the same approximation, as used by Davidson, Looser and Simmonds, we find that, at high frequencies,

$$A_\phi + A_\phi' = -\frac{2bnIs}{\pi\omega} \sin \omega t \int_{a-c}^{a+c} \int_0^{\pi/2} \frac{(Z-z)(2 \sin^2 \theta - 1)}{[(b+\rho)^2 + (Z-z)^2 - 4b\rho \sin^2 \theta]^{3/2}}\, d\theta dZ$$

and, accordingly,

$$i_\phi = \frac{2bnI}{\pi} \cos \omega t \int_{a-c}^{a+c} \int_0^{\pi/2} \frac{Z(2 \sin^2 \theta - 1)}{[(b+\rho)^2 + Z^2 - 4b\rho \sin^2 \theta]^{3/2}}\, d\theta dZ$$

By writing $k^2 = 4b\rho/[(b+\rho)^2 + Z^2]$, we obtain, with the authors,

$$i_\phi = -\frac{nI \cos \omega t}{4\pi b^{1/2} \rho^{3/2}} \int_{a-c}^{a+c} k(2K - \frac{2-k^2}{1-k^2} E)\, ZdZ$$

But $ZdZ = -2b\rho d(k^2)/k^4$

Therefore

$$i_\phi = \frac{nI \cos \omega t}{2\pi} \left(\frac{b}{\rho}\right)^{1/2} \int_{Z=a-c}^{Z=a+c} \frac{1}{k^3} \left[2K - (1 + \frac{1}{1-k^2})\, E\right] d(k^2)$$

Referring, now, to the integral formulae for complete elliptic constants, as given by Jahnke and Emde: "Tables of Functions" (Dover Publications—New York 1943) p. 78, and writing, also,

$u^2 = 4b\rho/[(b+\rho)^2 + (a+c)^2]$
and $v^2 = 4b\rho/[(b+\rho)^2 + (a-c)^2]$

we obtain, finally,

$$i_\phi = \frac{nI \cos \omega t}{\pi} \left(\frac{b}{\rho}\right)^{1/2} \left\{(1/u)\,[2E_u - (2-u^2)K_u] - (1/v)\,[2E_v - (2-v^2)K_v]\right\}$$

the subscripts denoting the moduli of K and E.

We may still simplify this expression by using the notation of Jahnke and Emde, and, actually,

$$i_\phi = \frac{nI \cos \omega t}{\pi} \left(\frac{b}{\rho}\right)^{1/2} (v^3 C_v - u^3 C_u)$$

The complete elliptic integral C appearing in the above final expression for the eddy current density is tabulated for intervals of 0.01 of k^2 on p. 82 of the above edition of Jahnke & Emde.

C. A. SIOCOS.

Stanford University,
California, U.S.A.

Beam Tetrodes

To the Editor, "Wireless Engineer"

SIR,—It is generally recognised that when the primary current density flowing into the screen-to-anode space falls below a limiting value, it is not then always possible to ensure the formation of a potential minimum in this space by the charges due to primary electrons alone; this situation is especially true when the anode voltage is approximately one-quarter of the screen voltage. It has consequently been suggested by several writers that the space charge arising from secondary electrons plays an important role in producing a potential minimum when the space charge due to primary electrons is inadequate. The computation of the magnitude of the effect is fairly difficult, and can only be performed if there is sufficient data regarding the amount of the possible emission and the velocity distribution.

A great simplification can be obtained, however, if we may postulate that the secondary electrons can be emitted copiously with zero velocities of emission; we then have the condition that the secondary electron flow is limited to the value which sets up just enough additional space charge in the

screen-to-anode space to reduce (dV/dx) at the anode to zero.

If this assumption is made the equation giving the potential distribution in the screen to anode space is

$$\frac{d^2V}{dx^2} = \frac{4\pi}{\sqrt{2\epsilon/m}}\left[\frac{I_P}{\sqrt{V}} + \frac{I_S}{\sqrt{V - V_2}}\right]$$

Where I_S is the secondary current density
I_P is the primary current density.

If I_S/I_P is denoted by γ, it can be shown that the following equation serves to determine γ :—

$$\sqrt{J_P} = \int_0^{(1-W_2)^{3/4}} \frac{dZ}{\sqrt{\frac{\sqrt{W_2 + Z^{4/3}} - W_2^{1/2}}{Z^{2/3}} + \gamma}}$$

Where

$$J_P = \frac{I_P}{I_D}; \quad I_D = \frac{\sqrt{2\epsilon/m}}{9\pi} \quad \frac{V_1^{3/2}}{x_a^2}; \quad W_2 = \frac{V_2}{V_1}$$

V_1 = screen voltage
V_2 = anode voltage
x_a = screen-to-anode gap.

I_D is, of course, the current density in a plane diode of anode voltage V_1 and gap x_a. The results show that for $J_P = 1$, γ does not exceed 0.2, even when $W_2 = \frac{1}{4}$. $J_P = 1$ may be considered to correspond to a current density less than half the peak current density, and the analysis confirms that the space charge due to secondary electrons does in fact considerably help to suppress the retrograde flow of secondary electrons.

S. RODDA.

Enfield, Middx.

Spectrum of a Phase- or Frequency-Modulated Wave

To the Editor, "Wireless Engineer"

SIR,—In a paper in *Wireless Engineer* for March 1944, Mr. F. M. Colebrook pointed out some interesting features of the spectrum of a wave sinusoidally modulated in phase or frequency. The object of the present note is to generalize these results somewhat and in particular to consider the case in which the carrier frequency is an integral or half-integral multiple of the modulation frequency. In this case each sidewave is a doublet and in general the mean square value of the wave is not equal to half the square of the amplitude.

A general representation of a wave of unit amplitude sinusoidally modulated in phase or frequency is

$$z = x + jy = \exp j\,[\omega_0 t + m \sin(pt + \phi)] \quad (1)$$

and by taking the real or imaginary part of z and by giving ϕ any value the complete range of cases can be considered.

The wave can be expanded into the spectrum

$$z = \exp(j\omega_0 t) \sum_{n=-\infty}^{\infty} J_n(m) \exp jn(pt + \phi) \quad .. \quad (2)$$

giving
$$x = \cos[\omega_0 t + m \sin(pt + \phi)] = \sum J_n \cos[(\omega_0 + np)t + n\phi] \quad .. \quad (3)$$

$$y = \sin[\omega_0 t + m \sin(pt + \phi)] = \sum J_n \sin[(\omega_0 + np)t + n\phi] \quad .. \quad (4)$$

Since n assumes all values from $-\infty$ to $+\infty$ it is seen that when n is negative and greater in magnitude than ω_0/p the frequency terms become negative but with the identities

$$\cos(-qt + n\phi) \equiv \cos(qt - n\phi)$$
$$\sin(-qt + n\phi) \equiv -\sin(qt - n\phi)$$

are transformed into the corresponding positive frequencies. This implies that there are two series of spectral lines :

(i) with $n \gg -\frac{\omega_0}{p}$ and (ii) with $n < -\frac{\omega_0}{p}$

which for convenience may be termed the positive series and the negative series respectively.

A special case is that in which the two series coincide in frequency which occurs when $2\omega_0/p$ is either an integer or a half-integer. If $2\omega_0/p = N$ it is seen that the nth and the $-(N + n)$th sidewaves coincide and have the frequency

$$\omega = \omega_0 + np = -[\omega_0 - (N + n)p] \quad .. \quad (5)$$

when N is even the lowest frequency is zero and corresponds to the single $(-\frac{1}{2}N$ th) component; when n is odd the lowest frequency is $p/2$ and is a doublet corresponding to the $-\frac{N-1}{2}$ th and $-\frac{N+1}{2}$ th components.

The spectrum (2) may then be written

$$z = \sum_{n_0}^{\infty} \{J_n \exp j(\omega t + n\phi) + J_{-N-n} \exp[-j(\omega t + \overline{N + n}\,\phi)]\} \quad .. \quad .. \quad (6)$$

where ω is related to n by equation (5) and n_0 is $-\frac{N-1}{2}$ when N is odd and $-N/2$ when N is even except that only *one* term of zero frequency is to be included in the summation.

The coincidence of the two series of sidewaves affects the calculation of the mean square value of the wave. When they are not coincident the mean square value is as Colebrook showed

$$\overline{x^2} = \overline{y^2} = \tfrac{1}{2}\sum_{-\infty}^{\infty} J_n^2 = \tfrac{1}{2} \quad .. \quad .. \quad .. \quad (7)$$

In the case of coincidence the mean square value may be greater or less than $\frac{1}{2}$ depending upon the values of N and ϕ. The value may be found either by summing the squares of the resultant side-wave amplitudes given by equation (6) or, more directly, averaging by integration over a period $2\pi/p$ of the modulation frequency. Since

$$\overline{x^2} + \overline{y^2} = |z|^2 = 1$$

it is only necessary to find x^2, and y^2 is then also determined. Now

$$\overline{x^2} = \frac{1}{2\pi}\int_0^{2\pi} \frac{1 + \cos[2Npt + 2m \sin(pt + \phi)]}{2}\, d(pt)$$
$$= \tfrac{1}{2}[1 + (-)^N J_N(2m) \cos N\phi] \quad .. \quad .. \quad (8)$$

by well-known integral formulae. The fractional departure of $\overline{x^2}$ or $\overline{y^2}$ from $\frac{1}{2}$ cannot exceed $J_N(2m)$ and in practice this will be an extremely small quantity since ω_0 $(= \frac{1}{2}Np)$ is always large compared with the maximum frequency excursion $\delta\omega$ $(= mp)$.

In any event it is doubtful if an ideal wave of the types represented by equations (2) and (3) can be very closely approached in practice. Nevertheless the influence of the ratio $\omega/_0 p$ on the spectrum and its content is of some intrinsic theoretical interest.

R. E. BURGESS.

National Physical Laboratory.

Pulse Modulation

To the Editor, "Wireless Engineer."

SIR,—Without committing ourselves to the detailed formulae quoted by Mr. Shepherd (*Wireless Engineer*, April, 1946, p. 114), we find ourselves agreeing with his general conclusions in point of rigour, but as he himself notes, the quantitative discrepancies are negligible in most practical cases. The contradiction which he mentions in connection with our Section 1.4.3 is no more real than that which might be adduced between any pair of first and second order approximations. That Mr. Shepherd's analysis is still an approximation seems to follow from his assumption that "r" is a constant, although it is not difficult to suggest how his method could be extended to more general modulation waveforms.

Other questions which arise in more exact analyses are whether in the case of length or delay modulation the trailing edge of the pulse is controlled by the modulation, and in the case of amplitude modulation whether the top of the pulse is truly flat or follows the modulation itself.

The analysis of length- and delay-modulated pulses with high modulation indices has already received attention in several quarters and will, no doubt, be published in due course. We look forward with interest to Mr. Shepherd's fuller account of his calculations.

F. F. ROBERTS.

London, N.W.2. J. C. SIMMONDS.

J. L. Baird

IT is with regret that we have to record the death of John Logie Baird at his home at Bexhill on 14th June after an illness which started in February. Born in 1888 at Helensburgh, Scotland, he was the son of the minister of the West Parish Church and was educated at Larchfield and the Royal Technical College, Glasgow. At the age of 18, when he was forced by indifferent health to give up London business life, he started a small television laboratory at Hastings.

By 1926 he had achieved true television and demonstrated the transmission of living subjects with light, shade and detail to members of the Royal Institution. In 1929 experimental television transmissions by his system were started, first by the German Post Office and then by the B.B.C. Shortly before the war Baird demonstrated colour television, using a cathode-ray tube and a series of colour filters to give a projected picture of 2ft 6in by 2ft. He has also the credit of carrying out the first transatlantic television demonstration in 1928 and of inventing the Noctovisor—an apparatus for seeing in the dark by invisible rays.

Since 1941 he has been consulting technical adviser to Cable & Wireless.

WIRELESS PATENTS

A Summary of Recently Accepted Specifications

The following abstracts are prepared, with the permission of the Controller of H.M. Stationery Office, from Specifications obtainable at the Patent Office, 25, Southampton Buildings, London, W.C.2, price 1/- each.

ACOUSTICS AND AUDIO-FREQUENCY CIRCUITS AND APPARATUS

574 036.—Preventing phase-distortion in loudspeaker installations when separate horn instruments are used to reproduce the higher and lower ranges of frequency.

Western Electric Co. Inc. Convention date (U.S.A.) 30th *December*, 1942.

574 370.—Loudspeaker system for projecting sound uniformly over a predetermined solid angle from one or more horns backed by an elliptical reflecting surface.

Western Electric Co. Inc. Convention date U.S.A.) 30th *December*, 1942.

574 394.—Acoustic horn of hyperbolic design which is stated to give optimum impedance conditions over a wide range of frequencies.

Jensen Radio Manufacturing Co. Convention date (U.S.A.), 23rd *July*, 1942.

574 463.—Construction and assembly of a multiplate piezo-electric transducer unit, say for driving a loudspeaker.

The Brush Development Co. (assignees of J. P. Arnot, Junr.). Convention date (U.S.A.) 11th *January*, 1943.

574 557.—Regulating the degree of automatic volume-expansion to be applied, when reproducing from any given sound record, by utilizing a photoelectric control-current derived from an auxiliary strip on the record.

The British Thomson-Houston Co. Ltd. and A. P. Castellain. Application dates 24th *August*, 1943, *and* 13th *January*, 1944.

574 611.—Directional microphone in which one side of the diaphragm is coupled to the atmosphere through a phase-shift network including a resistance and a chamber with spaced apertures.

Western Electric Co., Inc. Convention date (U.S.A.) 16th January, 1943.

DIRECTIONAL WIRELESS

574 319.—Radio-altimeter of the phase-difference type in which a cathode-ray tube indicates the presence of comparatively small reflecting obstacles projecting above the general level of the ground.

Standard Telephones and Cables Ltd. and R. M. Barnard. Application date 5th April, 1940.

574 340.—Radio course-indicating system of the overlapping beam type, in which the continued functioning of the transmitter is positively recorded even when the plane is on the correct " no signal " course.

Standard Telephones and Cables Ltd., C. W. Earp, and J. D. Weston. Application date 20th January, 1944.

574 546.—Electro-mechanical filter, involving commutation of the selected frequency, particularly applicable to direction-finding apparatus.

Radio Transmission Equipment Ltd. and C. E. G. Bailey. Application date 4th September, 1941.

574 610.—Radiolocation equipment in which exploring pulses are alternately switched, through wave-guides and reflecting mirrors, along two slightly divergent beams.

Western Electric Co., Inc. Convention date (U.S.A.) 19*th August,* 1942.

574 651.—Arrangement of spaced pick-up devices for indicating the location of any selected one of a number of sources which are radiating pulsed or sinusoidal waves of sound or other forms of energy.

Hazeltine Corporation (assignees of H. M. Lewis). Convention date (U.S.A.) 23rd May, 1942.

574 675.—Radiolocation system in which the exploring signals are radiated along a beam which is oscillating about a mean position, the critical direction being determined by the merging together of the reflected signals on a C.R. indicator.

Standard Telephones and Cables, Ltd. (assignees of H. G. Busignies). Convention date (U.S.A.) 8*th March,* 1941.

574 676.—Radiolocation system in which exploring pulses are radiated in rapid succession from the four limbs of crossed dipoles, and the position of a reflecting target is shown, either above or below, or to one side or other, of the centre of a C.R. indicator.

Standard Telephones and Cables, Ltd. (assignees of E. Labin). Convention date (U.S.A.) 13*th March,* 1941.

RECEIVING CIRCUITS AND APPARATUS

(*See also under Television*)

574 047.—Means for controlling or modifying the field-strength of a distant transmitter so as to diminish its interference effect at a given point of reception.

J. Robinson. Application date 2nd May, 1940.

574 075.—Preventing " mistracking " in the tuning control of a radio receiver, when the input inductance is partly shunted by a frame aerial.

Marconi's W.T. Co. Ltd. (assignees of W. F. Sands). Convention date (U.S.A.) 30*th December,* 1942.

574 078.—Cathode-ray receiver for making a visible record of incoming signals on an external strip of sensitized paper.

Marconi's W.T. Co. Ltd. (assignees of R. L. Snyder, Junr.). Convention date (U.S.A.) 31st *December,* 1942.

574 350.—Printing telegraph receiver of the facsimile type comprising a marking-wheel with flexible or pivoted arms for transferring the code signals to sensitized paper.

G. T. Evans, F. O. Morrell, and W. E. Finlason. Application date 17th January, 1941.

574 373.—Pocket-type receiver with a frame aerial which is normally housed in a recess in the casing, but which is hinged so that it can be used either as a carrying-handle, or as an easel-support.

Philco Radio and Television Corp. (assignees of G. Patterson, Junr.). Convention date (U.S.A.) 28*th July,* 1943.

574 407.—A frame-aerial feeding a permeability-tuned input-circuit and arranged to give constant-coupling over a wide frequency range.

Marconi's W.T. Co. Ltd. (assignees of G. L. Beers). Convention date (U.S.A.) 28*th January,* 1943.

TRANSMITTING CIRCUITS AND APPARATUS

(*See also under Television*)

574 204.—Method of connecting coaxial cables through a spacing-member which is formed with a cavity to hold a joint-sealing plastic composition.

Standard Telephones and Cables Ltd. and H. Birkby. Application date 15*th January,* 1944.

574 230.—Connecting two coaxial transmission-lines, so that whilst both are in axial alignment, one section of line can be rotated relatively to the other.

Bruno Patents Inc. (assignees of W. A. Bruno). Convention date (U.S.A.) 14*th November,* 1942.

574 349.—Printing telegraph transmitter of the facsimile type based on the use of a static keyboard and code contacts in the form of wires stretched beneath the keyboard.

G. T. Evans, F. O. Morrell, and W. E. Finlason. Application date 17th January, 1941.

SIGNALLING SYSTEMS OF DISTINCTIVE TYPE

574 627.—Repeater system for pulse-modulated signals, allowing simultaneous two-way working on one carrier-frequency, with a single aerial at each station.

Standard Telephones and Cables, Ltd. (communicated by International Standard Electric Corporation). Application date 11th February, 1944.

574 647.—Phase-modulating system in which a high degree of linearity is secured with the aid of a comparatively-low ratio of frequency-multiplication.

Hazeltine Corporation (assignees of B. D. Loughlin). Convention date (U.S.A.) 3*rd March,* 1942.

574 675.—Pulsed-signalling system in which the spacing between successive pulses is respectively decreased or increased in proportion to the value of the positive or negative signal-voltage.

Marconi's W.T. Co., Ltd. (assignees of C. W. Hansell). Convention date (U.S.A.) 29*th November,* 1940.

574 686.—Frequency-modulating system comprising a stabilized and a non-stabilized oscillator, and a mixing circuit which delivers a signal output derived from both oscillators.

Standard Telephones and Cables, Ltd. (communicated by International Standard Electric Corporation). Application date 11th February, 1944.

CONSTRUCTION OF ELECTRONIC-DISCHARGE DEVICES

574 430.—Process for reducing the brittleness of carburized tungsten filaments used in electron-discharge tubes.

Standard Telephones and Cables Ltd. (assignees of C. V. Litton). Convention date (U.S.A.) 23rd November, 1942.

574 453.—Means for controlling the cycloidal movements of the electrons in a discharge device of the magnetron type.

Standard Telephones and Cables Ltd. (communicated by Western Electric Co., Inc.). Application date 28th March, 1942.

574 480.—Spacing and supporting the electrodes of a discharge tube which is required to withstand high mechanical stress, or to be suitable for short-wave working.

Ferranti, Ltd., A. L. Chilcot, and S. Jackson. Application date 27th June, 1942.

574 500.—Process for making a metal-to-glass seal, suitable for withstanding high temperature-changes, in an electron-discharge device.

Philips Lamps Ltd., O. Pressel, and G. A. M. Diepstraten. Application date 3rd February, 1944.

574 512.—Discharge tube wherein the electron stream is focussed into the form of a ribbon and is velocity-modulated by electrodes spaced apart along its length.

L. F. Broadway and C. S. Bull. Application date 5th March, 1940.

574 551.—Short-wave magnetron comprising an annular resonator the interior surface of which is divided by a tortuous slot into two areas of opposite polarity.

"Patelhold" Patentverwertungs &c. A.G. Convention date (Switzerland) 9th February, 1942.

574 594.—Construction of valve in which two electrode-systems co-operate with different parts of the same elongated cathode.

The M-O Valve Co., Ltd. and C. W. Cosgrove. Application date 2nd February, 1944.

574 708.—Arrangement of a hollow resonator and an electron-discharge tube so as to minimize the dielectric losses caused by the glass envelope of the tube.

A. D. Blumlein. Application date 23rd September, 1940.

SUBSIDIARY APPARATUS AND MATERIALS

574 398.—Piezo-electric oscillator mounted between electrodes which are relatively rotatable to allow the operating frequency to be varied within limits.

The General Electric Co. Ltd. and S. K. Lewer. Application date 11th August, 1943.

574 401.—The use of "steel wool" as a screening or attenuating medium for preventing the leakage of centimetre waves through small apertures or passages.

The General Electric Co. Ltd., E. G. Janes, and A. O. E. Lindell. Application date 20th August, 1943.

574 428.—Current-converting system in which a "master" vibrating relay controls one or more "slave" relays in synchronism or in a given phase-relation.

H. M. Harmer and F. J. Cook. Application date 18th November, 1943.

574 501.—Assembling the metallized sheets, and fixing the terminal foils to a capacitor of the stack type.

M. Emanuel. Application date 4th February, 1944.

574 574.—Fixed capacitor consisting of a copper core, coated with vitreous material, and a superposed silver burnish, for use on aircraft, or military vehicles.

R. F. Oxley. Application date 26th June, 1942.

574 596.—Construction of a rotary-disc type of trimming capacitor to facilitate adjustment *in situ*.

Ashley Accessories, Ltd. and W. P. Kempster. Application date, 5th February, 1944.

574 612.—Design of ribbon cable for multiplying terminal banks in a telecommunication system.

Standard Telephones and Cables, Ltd. (assignees of G. Deakin). Convention date (U.S.A.) 5th February, 1943.

574 618.—Test equipment for indicating any deviation or eccentricity of the conductor, relatively to the dielectric, in an insulated cable.

V. Planer and F. E. Planer. Application date 9th February, 1944.

574 665.—Balanced-bridge circuit, comprising two adjustably-biased thermionic valves, for measuring either direct or alternating voltages.

Standard Telephones and Cables, Ltd. (assignees of J. R. Banker). Convention date (U.S.A.) 24th April, 1943.

574 775.—High-frequency ignition system, particularly for aircraft engines required to work at high altitudes.

The British Thomson-Houston Co., Ltd., D. F. Welch and A. Grieve. Application date 7th September, 1942.

574 816.—Thermo-electric generator comprising a battery of thermo-couples, with heating and cooling means to maintain a temperature-difference between the hot and cold junctions.

Eaton Manufacturing Co. Convention date (U.S.A.) 1st April, 1943.

574 830.—Valve-voltmeter circuit, with negative feed-back, for measuring either direct or alternating voltages.

Standard Telephones and Cables, Ltd. (assignees of J. R. Banker). Convention date (U.S.A.) 24th April, 1943.

574 837.—The use of a saturated "electret" (i.e. a dielectric in which static charges are permanently stored), as a frequency-changer, or as a modulating device.

Standard Telephones and Cables, Ltd. (assignees of G. C. Southworth). Convention date (U.S.A.) 25th March, 1941.